SYNTHESIS OF A BASIC RANGE CHANNEL FOR IMPLEMENTATION OF A COMPLETE MTI BY RANGE-GATED FILTERING

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THESIS

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June 1969

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for Implementation of a Complete

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Submitted in partial fulfillment of
requirements for the degree of

MASTER OF SCIENCE IN ELECTRICAL ENGINEERING

from the

NAVAL POSTGRADUATE SCHOOL
June 1969

ABSTRACT

A range-gated channel was constructed for use with various doppler filters in building the basic element of a radar Moving-Target-Indicator (MTI) by Range-Gated Filtering (RGF). Improvement over existing systems was accomplished by circuit simplification and solid-state design incorporating MOS devices in sampling circuits and d-c coupled amplifiers. Performance of the channel, using an R-C high-pass filter as the simplest doppler filter, was compared to that of the delay-line canceler MTI of the AN/UPS-1 air-search radar.

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SYMBOLS AND ABBREVIATIONS

f_c	Radar carrier frequency, cycles per second
f_d	Target doppler-frequency shift
f_r	Pulse repetition frequency
PRF	Pulse repetition frequency
n	Number of pulses returned from a single sweep of radar antenna pattern beamwidth across target
λ	Radar wavelength, in centimeters
$\dot{\theta}$	Radar antenna rotation rate
θ_B	Radar antenna pattern horizontal 3db beamwidth
T	Pulse repetition interval
v	Volt
v_r	Target relative velocity
MOSFET	Metal-Oxide-Semiconductor Field-Effect Transistor

I. INTRODUCTION

Moving-Target-Indication is a mode of radar operation in which doppler shift of the carrier frequency, caused by any reflector having a finite relative-velocity vector normal to the plane of the radar antenna, is used to detect moving targets and to eliminate the deleterious effects of echoes from fixed targets which are not of interest. Operationally this enables a radar operator to discriminate between echoes from moving targets and return from environmental reflectors such as terrain, weather, and ocean waves. These "fixed-target" echoes would obscure the moving target of interest in the normal mode of search radar operation. On a visual display such unwanted returns are termed "clutter," because they do, in effect, clutter the display.

Inspection of the formula relating reflector relative velocity to the resultant doppler frequency

$$f_d = \frac{103\ v_r}{\lambda}$$

indicates a simple, direct relationship. By their nature clutter-producing reflectors normally have low relative velocities, whereas moving targets, such as aircraft, have much higher relative velocities. It follows immediately that frequency filtration could be used to eliminate low-frequency clutter while preserving returns from higher-frequency targets. There are three complicating factors, however.

First, rotation of the antenna introduces a modulation component for all signals, the effect of which is a spreading of the clutter spectrum by an amount proportional to the speed of rotation. This necessitates a higher cutoff frequency to insure complete clutter rejection.

Second, recent studies [1] have indicated that the returns of certain clutter produce much higher frequencies than might have been expected. The implication is that if the lower cutoff-frequency were extended high enough to eliminate all clutter, returns from slower-moving aircraft (for the air-search radar case) might be eliminated. Hence, a highly desirable feature of an MTI system is a variable low-frequency cutoff, the actual value of which would be determined by existing conditions during operation and anticipated target velocities.

Examination of the frequency response of a typical pulse-type search radar reveals the third, actually prohibitive obstacle to the use of this simple solution. A typical frequency spectrum, including clutter and a target, is shown in Figure 1 (receiver noise is ignored). The clutter spectrum is reflected about the pulse repetition frequency (PRF) f_r, and its multiples as shown. For coherent detection of targets the target appears in the same relative position in each PRF interval. Obviously, filtration of only the lowest frequencies about the carrier frequency would eliminate only a small portion of the total clutter signal. Rather, a comb filter having notches at f_c and other frequencies separated from it by multiples of f_r, as suggested by the dashed line, is required for complete clutter rejection.

Until recently the operational solution for obtaining this comb filter has been the delay-line canceler. In this method one or more delay lines may be used to synthesize a wide range of filter characteristics, operating on a video signal which has been mixed with a suitable carrier frequency. The advantages of this technique has always been simplicity of implementation and relatively small size. However, once a particular delay line has been built, its delay time is fixed and its use is therefore restricted to radars having a PRF equal to the reciprocal of the delay time. Additionally, the effects of the intrinsic characteristics of delay-line materials, usually quartz or mercury, result in several disadvantages. [2] They include: large insertion loss; tight tolerances on phase-shift characteristics which must be precisely matched in parallel channels; secondary signals arising within the delay line from several internal sources; extreme temperature dependence; and the necessity for maintaining constant pulse amplitude and stable pulse repetition frequencies. Any imbalances or inaccuracies result in degradation of performance, and this degradation can be severe for certain configurations. For a discussion of single and double delay-line canceler performance, see Reference 3.

Range-gated filtering (RGF) is another method of implementing MTI. While it offers several advantages over delay-line methods, its main disadvantage has been circuit complexity, resulting in large size and high power consumption. In the past this consideration alone has always prevented actual system implementation. Advances in semiconductor technology and microminiaturization have removed this obstacle, and poor results from present systems operating in high-clutter environments in Southeast Asia have prompted new investigation of MTI by RGF and attempts to synthesize a high-performance, practical system.

Several such systems have actually been built and tested. References 4 and 5 provide examples of systems designed for respective application in non-coherent and coherent radar receivers. Reference 6 provides an example of the degree of miniaturization which has been achieved. Although a variety of circuitry has been employed, the basic operation and results are the same. Accurate timing for RGF systems is essential, but this demand seems to have been adequately met. However, potential for further refinement of the range channels themselves does exist.

II. OBJECTIVES

The objective of this work was to synthesize a single range channel which could then be used in conjunction with various doppler filters in designing a complete MTI by RGF. Designing for future implementation in microminiaturized form, all solid-state components were used. Within the context of specific circuit operations, existing methods were compared and further simplification or improvement made where possible.

A very desirable feature of such a system would be the capability of operation without continuous check and adjustment by an operator, especially in the military environment. However, as previously explained, it is also very desirable to have the capability of operator adjustment of the clutter filter characteristics, as well as control of the threshold level for target detection. These considerations also served as guidelines for design.

The AN/UPS-1 air search radar was chosen as representative of its class, and it was therefore used for comparison of MTI performance. This radar, employing a coherent detector and delay-line canceler MTI, operates with a PRF of 800 cps, pulse width of 1.4 microseconds, and antenna pattern horizontal 3db beamwidth of 2.5 degrees. A typical antenna rotation rate of five rpm is used. The subject system was designed to be compatible with these characteristics; however, an advantage of the RGF method over delay-line cancelers is the ease of adaptability to radars with other timing schemes, and this capability was preserved. In order to provide a standard for improvement, a worst-case doppler filter consisting of a simple high-pass R-C circuit was inserted and performance compared to that of the AN/UPS-1 MTI. The method by which a single high-pass filter can effect clutter filtration is dependent on the operating characteristics of the preceding sample-and-hold circuit. This is explained in the following section.

III. BASIC OPERATION OF MTI BY RANGE-GATED FILTERING

The basic operation of RGF is simple and can be understood with the aid of Figure 2. The radar return is detected at IF by a detector designed to preserve the doppler frequency shifts, a phase detector in the AN/UPS-1, and the video signal is fed to the common input terminals of the parallel range channels. The timing circuit is activated by the system trigger, indicating transmission of a pulse, and this circuit then sequentially triggers the sampling gates to sample the video for an interval approximately equal to the radar pulse width. Therefore the input that each channel receives represents the radar return from a particular range interval. It is obvious that each channel will process only the target signals, clutter, and noise that are sampled for that short interval. This system thus eliminates collapsing loss due to a too-narrow video bandwidth. It is also obvious that if during the following filtering process the range information is destroyed, the range resolution of the system will be determined by the width of the sampling gate. A wide pre-detection bandwidth is assumed to maintain range information. Since narrow-band doppler filtering is actually used in the system built, this turns out to be the case.

The sampled signal is placed on a holding capacitor to be held for one pulse repetition interval, after which a new sample is applied. It is a characteristic of sample-and-hold circuits that input signals equal to any multiple of the sampling frequency itself appear in the output as d-c values, and the highest fundamental frequency which will appear in the output is one-half the sampling frequency. Figure 3 illustrates the first characteristic. If samples are taken at times 1, 2, and 3, and are held for one complete cycle of signal A, the output will be a d-c voltage of amplitude -S. Signal B is exactly twice the frequency of A, yet it is obvious that sampling it at frequency A will result in a zero d-c output. In the same manner, any sampled frequency which is a whole multiple of the sampling frequency will produce a d-c output whose value depends on the phase of the signal at the instant it is sampled. If the output is capacitively coupled, then all such signals produce zero output.

As the signal frequency increases and surpasses the sampling frequency, the situation shown in Figure 4 exists. Here the sample is held for more than one cycle of the signal, and the actual output appears as a much lower frequency. Further signal frequency increase leads to the situation shown in Figure 5, in which samples are taken at opposite peak values. It can be seen that the output appears at a frequency equal to exactly one-half the sampling rate, and for further signal frequency increase the output frequency will actually decrease until a d-c value appears at the next multiple of the sampling frequency.

The effect of the operation of the sample-and-hold circuit is therefore twofold with respect to frequency. First, it converts all signals at whole multiples of the sampling frequency into d-c voltages. These d-c voltages are eliminated by the high-pass doppler filter. Such filtering also introduces the problem of blind speeds, in which a moving target may produce a doppler frequency at a multiple of the PRF and be completely eliminated by the MTI itself. This problem is also characteristic of delay-line-canceler MTI operation.

The second result of this circuit is that all doppler signal frequencies appear in the output, but converted to a maximum frequency of one-half the sampling frequency. This excludes multiples of f_r, which appear as d-c and are filtered out as previously explained. Examination of the response to a frequency range lying between two multiples of the sampling rate (PRF) shows that application of high-pass filtering at the sample-and-hold output will yield, in fact, a comb filter having notches at multiples of the PRF. Choosing the low-frequency cutoff so as to provide the required clutter filtration then yields the exact solution which Figure 1 shows to be necessary. After the sample-and-hold circuit, only the frequencies between the lower cutoff-frequency and one-half the PRF are necessary for complete target information.

The obvious next step is doppler filtration, in which the low-frequency cutoff is established. Since frequencies greater than one-half the PRF are now redundant in their information content, they may be eliminated in a simple low-pass filter whose 3db frequency is greater than one-half the PRF. The important feature of the doppler filter -- which has been the most difficult to obtain -- is its low-frequency characteristic.

Figure 6 illustrates the state of the art in synthesizing suitable low-frequency rejection characteristics. Ideally the low-frequency cutoff would occur at the edge of the clutter spectrum, with a sharp rise in the gain characteristic for frequencies just above cutoff. The exact value of the cutoff frequency should be adjustable in accordance with clutter conditions. Such a characteristic is necessary to prevent unintentional elimination of targets in the edge of the clutter spectrum. The dash-dot line indicates what has been obtained by circuits which can be useful in RGF applications. Much work has been devoted to the area of doppler-filtration, and many references may be consulted in the literature. Of particular interest are the characteristics of canonical configurations, whose complexity and size problems might also be solved with microminiaturization. Reference 7 provides results of calculations for and implementation of such a filter. In view of the many filters available, it was decided to design the channel for insertion of already-synthesized units.

Following filtration the signal-to-noise ratio may be improved by non-coherent integration, which is effectively a summation of the target pulses. A simple technique which is often used is a simple R-C low-pass circuit. It has also been stated that the signal-to-clutter ratio for spatially-uniform clutter can be improved by the same method. [3]

In discussion of signal integration, Skolnik has stated that the conversion of signal energy to noise energy during rectification before integration causes such post-detection integration to be less efficient than pre-detection integration. [2] This is, therefore, an area to be considered for improvement.

The output of the integrator is a d-c voltage, ideally occurring only if a target is present within the range interval covered by that channel. In order to perform video reconstruction the integrator outputs can then be combined in an OR circuit. For the purpose of preventing apparent target detection due to noise, a signal threshold level can be applied before or after video reconstruction. Here again a simple solution was sought.

Proper interfacing for visual display completes the basic MTI by RGF.

IV. RANGE CHANNEL DESIGN AND OPERATION

Employment of the synthesized RGF channel in the AN/UPS-1 radar required use of the radar PRF, pulse width, and beamwidth in determination of certain component values as will be explained. In order to make a valid comparison of MTI performances, the detected video signal was taken from the test point available just prior to signal entrance into the MTI unit. The signal at this point is bipolar, having an amplitude of $\pm$ 0.6 volts. Operation of the radar MTI and the synthesized channel on this identical signal was therefore parallel, and the comparison of outputs was therefore considered valid. For the purpose of closer inspection of the range-channel output, the last stage of the channel was designed merely to provide input to a high-input-impedance oscilloscope, the need for addition of simple low-output-impedance circuitry to supply a radar visual display being recognized.

Figure 7 shows a block diagram of the entire circuit as synthesized, and Figure 8 is the circuit schematic diagram.

A. VIDEO DRIVER

Although the first actual stage of the range channel proper is a sample-and-hold circuit, in order to charge and discharge the holding capacitor to the full amplitude of signal excursion large current drive and low-impedance path to ground are necessary. Such demands can be fully appreciated by noting that, for a 400 cps signal (or whole multiple thereof), any two consecutive samples may have maximum values of opposite sign. For this case the voltage on the capacitor must change by twice the maximum signal amplitude. Provision for supplying and withdrawing the requisite charge must therefore be made. The Darlington-pair used in an emitter-follower configuration proved to be very adequate for this purpose.

As can be seen in the circuit schematic of Figure 8, the possibility exists for d-c voltage buildup on the combined input capacitances of the parallel sampling gates. These capacitances are shown by the dashed-line connections. Direct coupling from the video driver to the common input point was therefore used to prevent this, as well as to set the a-c signal at a desired d-c base level. This was required to allow proper operation of the sampling gate configuration which was used, as will be explained below.

B. SAMPLE-AND-HOLD CIRCUIT

The basic purpose of the sample-and-hold circuit at the input to each range channel is to separate the return signal into increments of range (time) and thus permit narrow-band filtration of the signal without loss of range information. The range resolution is thus equal, at best, to the width of the sampling pulse. For this case the radar pulse width equals the sample width, and the 1.4 microsecond channel coverage means a resolution of 210 meters.

Proper operation of the sampling gate itself requires an input time constant which is much less than the sample width, thus allowing the capacitor to fully charge or discharge during the sampling period. This means that a gate with very low on resistance is required. A small hold capacitor is desirable to minimize the time constant.

Once the sample has been placed on the capacitor, it must be held for a complete pulse repetition period. This requires that leakage paths back through the gate and forward through the following circuitry be of extremely high resistance. Two important criteria for an acceptable gate are thus a very low on resistance and a very high off resistance.

Several other important criteria were considered in choosing a gate. A very desirable feature of sampling gate operation is zero offset voltage, i.e., no appearance of the gating pulse itself in the output as a pedestal for the signal to "ride on." The total effect of equal pedestals after video reconstruction of the outputs of all range channels would be a d-c voltage, and this can easily be filtered out. However, spikes due to differentiation of the pulse edges could appear, so elimination of this pulse coupling was considered desirable.

The design goal of small size and weight precluded the use of any transformer coupling in either the signal or gating paths, and this became a limitation on several possible diode and transistor configurations. Partially related, another design decision was to specify provision of a gate-pulse source in each channel. This would greatly simplify the triggering and timing system itself, for only a small pulse would be necessary to trigger a source such as a monostable multivibrator. Since the need for a large gating pulse of 15 to 30 volts was anticipated, this would also help eliminate the undesirable coupling between timing and signal paths.

A very important characteristic of a gate operating with short sampling times is high speed, with switching times in the low nanosecond range necessary.

Economy of power as well as space must be considered for this application. Each channel incorporates two gate circuits, and the necessity for simplicity and economy can be appreciated by noting the number of them required for a complete MTI. For example, using a 1.4 microsecond sampling width for a 40 nmi range would require 353 channels, thus requiring 706 gates.

Four types of gates were considered, and tradeoffs between the various desirable characteristics were necessary. Diode configurations are widely used in sampling circuits and have an advantage of low on resistance and high off resistance. Bipolar transistors may be used to take advantage of their low on resistance. The JFET, with additional circuitry, has proved to be very useful. Reference 7 provides a summary of pertinent features of these types of gates.

Investigation of the qualities of the MOSFET [8, 9, 10] resulted in the decision to build a circuit with this device and to determine if its excellent potential was realizable. Recent development of devices having low channel on resistance has virtually eliminated this former drawback, and typical values of 30 to 40 ohms may be realized. In other respects the MOSFET appears to have great advantage over all other types.

A configuration was chosen in which the drain-source channel was used, in effect, as a variable resistor, with the gating pulse being applied to the gate. In this manner the low on resistance is used during sampling, and during the hold period the typical off resistance of 10^{11} ohms results in little leakage from the hold capacitor. [11]

While the MOSFET is not superior with respect to switching times, turn-on and rise times less than 10 ns may be expected, while turn-off and fall times of approximately 50 and 200 ns, respectively, may be obtained. Increasing the drain-source current will decrease the turn-off and fall times considerably.

Since the gating-pulse and signal circuits are entirely separate, zero offset voltage is theoretically possible. This was not obtained in this configuration, however, because the capacitance of the device itself allowed partial coupling between the two circuits. In this configuration the amount of coupling is directly proportional to the ratio of gate-signal circuit capacitance to capacitance between the signal circuit and the reference level. Thus the comparatively large hold capacitor and the small input capacitance of the following component combine to practically eliminate this effect in the input gate. The effect on output gate operation is discussed in a following section.

From an economy point of view the MOSFET gate is far superior to other devices. In order to pass the bipolar video signal optimally, i.e., by meeting bias requirements, the signal was set on a negative d-c level. This could be done by a-c coupling the signal to a large resistor returned to the bias voltage, but the method employed was chosen not only to satisfy this requirement, but also because it allowed optimal sample-and-hold operation. Including a resistor to return the gate to ground during the hold period, (a zero or positive gate-drain voltage completely turns off the device), the entire gate is obtained with two components. For best results the gate should be driven very negative with respect to the drain and source, and thus the requirement for a fairly large gating pulse.

With respect to power drain the MOSFET is very nearly ideal. The gate is connected to ground to turn the device off, and there exists a practically infinite resistance between the signal path and ground through the device. Only leakage currents in the sub-nanoampere range occur, meaning practically zero power consumption. The gate requires only a capacitively-coupled voltage pulse to turn the device on, and it provides negligible load to the pulse source. Thus no form of complicated coupling and pulse-source circuitry is required in either the signal or gate-pulse circuits.

The last stage of the sample-and-hold circuit is a high-input-impedance device, which is necessary to isolate the hold capacitor from ground and also to transmit the sampled signal. Here again the MOSFET was used to advantage. While the d-c bias superimposed on the signal was an aid in optimal transmission of the signal through the gate, its primary design purpose was to provide required negative bias for the input of this MOSFET. The similarity of operation of this device to that of a vacuum tube is apparent in this application. The very large input impedance, in excess of 10^{13} ohms, means essentially no leakage path is provided through which the capacitor voltage can decay. A simple MOSFET-bipolar combination was used to allow a small amount of gain before losses in filtering. [13] This amplifier was designed for modest gain to suit the needs of possible filter arrangements in the following stage. Due to the large amount of negative feedback from the source and emitter resistors, this is a very stable configuration. For the simple R-C filter employed in testing, a gain of two was used.

C. DOPPLER FILTER

The purpose and characteristics of the doppler filter unit of the channel has been explained, and many solutions have been formulated and constructed. The range channel was designed to allow insertion of arbitrary filters for testing. Therefore, in order to establish a reference level of performance for later comparison, a simple R-C high-pass filter was inserted and its performance investigated.

D. FULL-WAVE RECTIFIER

Following doppler filtration the signal is full-wave rectified before integration and video reconstruction. In anticipation of the use of diodes in the rectification process, with attendant introduction of noise and degradation of the integration efficiency, amplification at this point would seem logical. As will be seen, however, these problems were separated and solved separately. Some amplification will be required here for certain high-loss filters, however. Any stable amplifier with adjustable gain could be used; a single MOSFET stage was found to be sufficient for the filter employed here.

The method of phase splitting was employed to provide two equal but opposite-phase outputs. Each output was a-c coupled to a large load resistor, from which respective positive outputs were taken through diodes. The diode cathodes were then connected directly, without a load. The result is a voltage output in which no current flows through either diode, and thus no diode noise is introduced. A second advantage of this configuration is the elimination of the threshold effect which current-carrying diodes would introduce. In order to take advantage of this diode circuit, however, the next stage must have a very high input impedance.

Again the characteristics of the MOSFET lend themselves to advantageous use in the phase-splitter and high-input-impedance circuits. Since the drain and source currents of the MOSFET are theoretically equal, placement of equal load resistors in the respective terminal circuits should give equal a-c outputs of opposite phase. With closely matched load resistors and a separate gate-biasing circuit, this method proved very acceptable. Both outputs are follower circuits, in effect, and the gain for each output is therefore less than unity. The maximum input voltage of +1.6 v for the following amplifier determined that an input of $\pm$ 2.4 v to the phase splitter be used, and the simple one-stage MOSFET amplifier provided this.

The very high input impedance of the MOSFET resulted in its use in the amplifier which receives the signal voltage from the rectifier output. Bias problems were solved by connecting the source to a positive supply, thus eliminating connection of any components to the input terminal and affecting operation of the diodes. Regardless of conditions at the drain or source terminals, the input will be set at a zero volts reference level due to the small leakage which exists.

A very large resistor could be used to be certain that d-c buildup on the input capacitance does not occur. With the diode configuration used all voltages at the input will be greater than zero.

E. INTEGRATOR AND VIDEO RECONSTRUCTION

In designing the integration and video reconstruction circuits, full consideration was given to simplicity and to the goal of incorporating a single threshold for controlling the outputs of all range channels in the complete system. A very successful result was obtained by using d-c coupling of the output to provide biasing for following stages as well as to preserve the d-c output of the integrator. This is necessary because the integrator gives an essentially d-c output for targets, and a-c coupling would necessitate use of diodes to level set the signal. The final design eliminated the need for diodes and a-c coupling.

In anticipation of the signal attenuation caused by the integration process, amplification of the rectified signal was desirable. Inspection of the operating curves of the single MOSFET already employed showed that this device alone could not only solve the "diode load" problem already explained, but it would also provide all necessary amplification. Furthermore, its d-c output level could be so adjusted that biasing for a simple video-reconstruction circuit was obtained. This amplifier is thus the key to operation of both the integration and video-reconstruction processes. Figure 9 shows the operating line on the characteristic curves. A gain of more than 12 is obtained for an input excursion of +1.6v. Use of the positive portions of the rectified signal is now easily understood. Curves for both upper and lower temperature extremes show that this particular mode of operation is relatively stable with changes in temperature, and proper operation will still be obtained over the full military temperature range. The threshold voltage variation between devices could be a source of difficulty if discrete devices were used in each range channel, a problem which would be essentially eliminated by integrating all on the same chip.

Placement of a variable resistor in the source circuit allowed setting the d-c output level at zero, which allowed use of the same biasing scheme in the OR amplifier in the video reconstruction process. A MOSFET in a source-follower configuration was used with a positive source supply, with the input being returned to ground through a small resistor. The output voltages of the respective channels are developed across this resistor, and it also serves to prevent d-c buildup on the MOSFET input capacitance and the combined output capacitances of the gates. The need for current drive as in the input sample and hold does not exist, and the small load resistor was sufficient for complete transmission of the output signal.

The simple MOSFET sampling circuit was again used in the output of the channel. Since the signal to be transmitted during the gating period is a small d-c value below threshold voltage, returning the gate to ground is sufficient to open the signal circuit, and a -24v pulse is sufficient to completely turn on the gate for transmission of the sample. Although there is more coupling of the gating pulse to the signal path for a single gate, the parallel effect of the output capacitances of the other gates (which are turned off) greatly reduces this effect. Since all gates produce the same coupled output, the effect is production of a d-c voltage which will be eliminated by capacitive coupling to visual display circuits. Also, since the signal itself is negative, use of a diode to level set the signal will essentially eliminate the problem of positive spikes in the output due to differentiation of the trailing edge of the gating pulse.

The value of integrating time constant, RC, was determined by use of the formula

$$n \left(\frac{T}{RC}\right) = 1.257$$

which provides for the highest efficiency of exponential RC weighting in comparison to the ideal method of uniform pulse weighting. [2] Having assumed an antenna rotation of rate of five rpm, the formula

$$n = \frac{\Theta_B f_r}{\dot{\Theta}}$$

gives n equal to 66 pulses received from a pass of the beamwidth across the target. The optimum value of RC is therefore approximately 65 ms. By use of a large resistor in the integrator circuit, both the charge and discharge time constants were set close to this value.

A simple method of setting a threshold is by use of variable biasing, which can be implemented in the last stage. For the MOSFET used in the final stage of this circuit the signal was set on an arbitrary d-c level -- zero volts due to the small input-circuit resistor. This level can be effectively changed by adding dc to the signal, due to the manner of voltage biasing employed in the MOSFET. By placing a variable resistor in the source circuit of the pre-integration amplifier this d-c level can be varied. This is possible because d-c coupling to the final stage was used.

This method of setting the threshold also helps reduce the problem of drift in the d-c level of the output of this amplifier, because operator adjustment of threshold will automatically account for this. For the single channel that was constructed this adjustment also proved sufficient to completely cancel the portion of the gate pulse that was coupled into the signal circuit. Thus a visually-aided adjustment at this single point performs three adjustments in one, with the advantage of simplicity for the operator. Since this adjustment is in effect a change of positive potential at the source terminal, this single adjustment can be used for all channels of a complete MTI.

F. RADAR-COMPARISON TEST SET-UP

The set-up employed to determine the performance of the synthesized range-channel is shown in Figure 10. The operation of the two MTI systems is in parallel, with a moving target being simulated by the signal generator. Although a solid-state unit would be employed as the gate-pulse source in an actual system, the simple blocking oscillator was used as a matter of convenience. Its pulse width is variable between .9 and two microseconds, and its output was found to be more acceptable than that available from the pulse generator at hand. The signal generator itself acts as a delayed trigger for the blocking oscillator. Triggered by the system trigger, it provides a negative pulse whose trailing edge was differentiated and set to trigger the blocking oscillator. By this method, with the amplitude of the injected target being reduced to essentially zero, the channel range interval could be set to cover any desired 1.4 microsecond interval of the total range.

V. RESULTS

The range channel was constructed and incorporated into the test set-up shown. Because of the adjustable triggering system, this is effectively a complete radar system having MTI by RGF. However the system is complete for only the single range interval covered by the channel.

The primary result sought was the successful performance of the channel in conjunction with the radar. This goal was fully accomplished. All components and units performed as well with the radar as on the bench.

The purpose of the RGF is two-fold if stated in general terms -- preservation of range information while permitting detection of targets submerged in clutter. The integrator is also important in improving the signal-to-noise ratio. Two measures of the channel performance which provide an excellent means of comparison with the radar delay-line MTI are the minimum discernible signal and sub-clutter visibility. Minimum discernible signal determines the weakest signal that the receiver can detect, and reflects any improvement in the signal-to-noise ratio effected by the unit. Sub-clutter visibility indicates the increase in signal-to-clutter energy effected by the MTI, and this is the best single criterion for comparison. Since both MTI systems process the same signal from the radar, this comparison is fully valid. This would not necessarily be so for a comparison of different radars, for their resolution cells might not be equal, and therefore the reflected clutter power would be different.

Both measurements were made for each system, employing an injected moving target of known amplitude, and observation of the respective video signals on the CRT. Values were obtained by summing the coupling losses with the setting of the variable attenuator of the signal generator. Due to the particular radar environment of the AN/UPS-1 used, the determination of sub-clutter visibility was difficult and not highly accurate. However by using the same reference conditions (in accordance with standard operator procedures for finding this value), a comparison of the respective values should be valid.

For the delay-line canceler MTI of the radar a minimum discernible signal of -110 dbm was measured, which is approximately the reference value that was expected. The constructed channel allowed target detection at approximately -118 dbm, and at -115 dbm the target was more than evident. For the purpose of making observation of the channel output more accurate, the threshold was set so that the noise, which is a negative signal just as is the target, appears as a small a-c signal. This is actually setting the threshold at half the amplitude of the peak noise signal, but due to operation on both sides of the Q point of the final-stage OR amplifier, the output contains the positive output as seen. Elimination of the positive portion would indicate the actual output, with the threshold at zero level.

Measurement of sub-clutter visibility gave approximately 20 db for the delay-line canceler, which is considerably less than the reference value of 26 db. The RGF channel showed a sub-clutter visibility of between 20 and 21 db, being a little better than the delay-line canceler.

Figure 12 gives a comparison of outputs for an injected moving target with strength of -107 dbm. The top exposure is the delay-line output for the target seen in the bipolar video return of the second exposure. For this signal the doppler signal is easily discerned in the video signal. The third exposure is the output of the RGF channel, which was triggered so as to process the range interval containing the target. This output may be compared to the normal, noise-only, output shown in the lower exposure. Figure 13 shows the same four signals, but the signal strength is reduced to −113 dbm. The target is barely visible, even after close examination, but the RGF channel output convincingly proves its presence.

For the purpose of determining the effect of the channel itself on low-frequency response, aside from doppler-filter characteristics, large capacitors were substituted for the coupling capacitors in stages following the doppler filter. The difference may be noted in Figure 14, where a signal strength of −107 dbm was set. The top exposure is the reference output for noise only. The second exposure is clearly a doppler echo within noise. The third exposure shows the output with the large coupling capacitors inserted, while use of the smaller capacitors gives the output shown in the lower exposure. The channel clearly produces low-frequency filtration. Measurement of the minimum discernible signal with the large capacitors showed an approximate three db sensitivity decrease. The channel itself appears as a two-stage circuit having R-C coupling. Therefore it must be carefully designed so that the combined low-frequency response of the doppler filter and the channel determines the overall low-frequency characteristic.

Two areas of special interest are the amount of droop, or decay, which occurs in the sample-and-hold circuit, and the frequency response of the channel alone. A comparison between its characteristics and the desirable comb filter characteristic is important. Both areas were separately investigated, and the results for both were very good. With a maximum sensitivity of 10 millivolts, no droop of the value stored on the capacitor was discernible. This is a very desirable result, for sensitivity limitations would be the result of signal decay. This result is much better than the three per cent which some systems have tolerated.

Figure 15 shows the frequency response of the channel alone, without a doppler filter, for a frequency range of zero to 1600 cps. The response was checked to 40,000 cps and found to duplicate periodically the response between 800 and 1600 cps, as expected. The lower response for frequencies up to 800 cps is due to low-frequency characteristics of the video driver and sample-and-hold circuits, and this does not affect low-frequency response after the buffer amplifier.

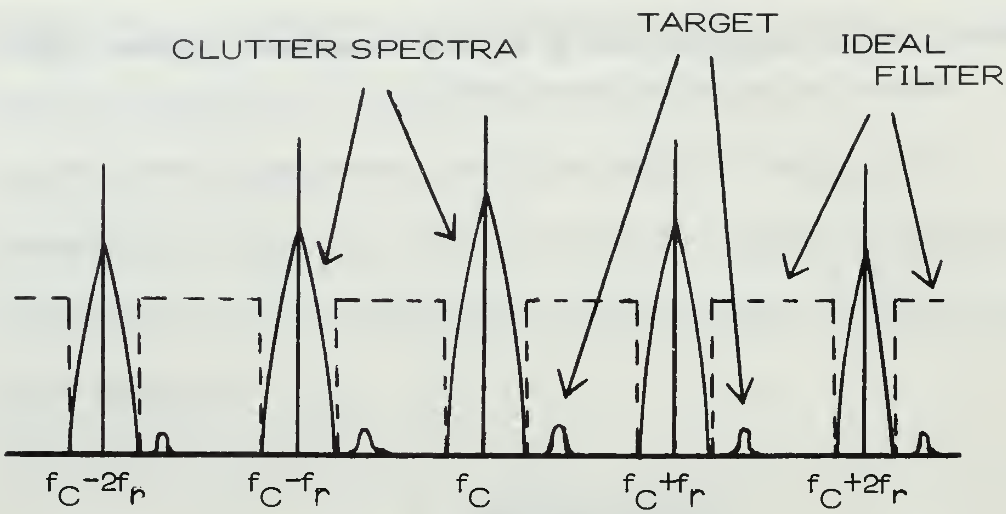
Figure 16 illustrates the sampler output for a frequency of 400 cps. This is a critical area with respect to sample-and-hold operation as previously explained, and a decrease in gain is expected. The small amount of this decrease is indicative of very good sample-and-hold operation.

Figure 17 illustrates output at the same point for a 110 cps input signal, which appears as expected.

As a matter of general comparison, Figure 18 gives a comparison of outputs for an aircraft at a range of 20 miles (the channel trigger was adjusted to set the channel on the target). The first exposure is the radar MTI output for the large doppler signal seen in the second exposure. (The large V is a fixed target return, which is rejected by both MTI systems.) The middle exposure shows the channel output and the same target return, with the target not being within the coverage of the channel. When the channel trigger is adjusted for target detection, the output of the channel appears as shown in the lower exposure.

VI. CONCLUSIONS

As pointed out previously, several complete MTI systems employing range-gated filtering have been successfully built and tested. The purpose of this work was to simplify and improve the basic element of the RGF -- the range channel -- and incidentally to provide a means for testing various doppler filters for such a system. Comparison of the design with existing designs shows that significant simplification has been accomplished, and that improvement has been made in several areas, notably in gate and sample-and-hold circuitry. Comparison of performance with that of the AN/UPS-1 shows this system to perform as well, and probably better. This was accomplished with the simplest of doppler filters. Thus the results were considered to be very successful, and to be further proof of the simplicity and effectiveness of this type of MTI. While further improvement is of course necessary, especially in the stability of the last stages, a very acceptable system might be obtained immediately by a small improvement in the channel low-frequency characteristics and insertion of a more acceptable doppler filter.



AMPLITUDE VS. FREQUENCY

FIGURE 1. PULSE-RADAR FREQUENCY SPECTRUM

FIGURE 2. BLOCK DIAGRAM OF MTI BY RANGE-GATED FILTER

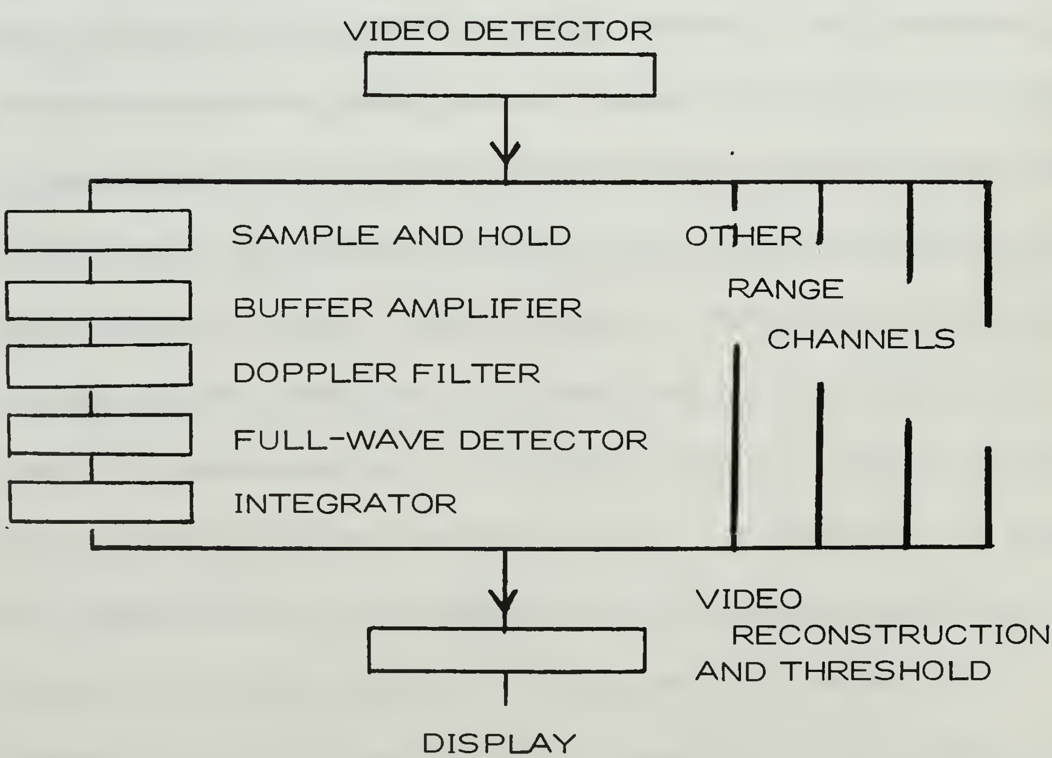


FIGURE 3. SAMPLE-AND-HOLD OUTPUT FOR MULTIPLES OF THE PRF

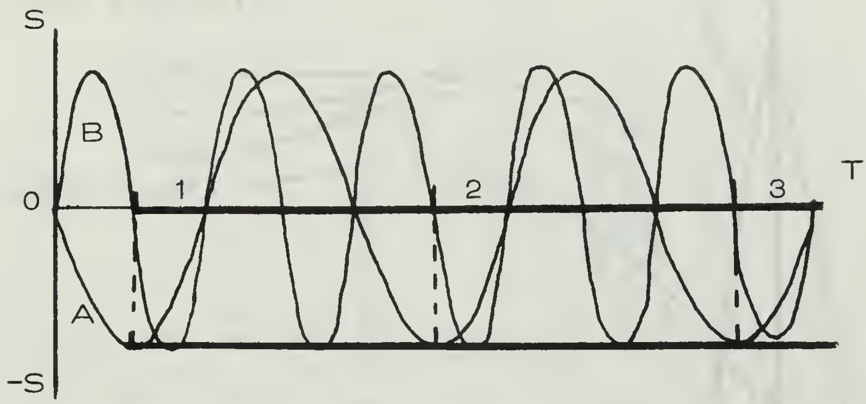


FIGURE 4. SAMPLE-AND-HOLD OUTPUT FOR AN ARBITRARY FREQUENCY

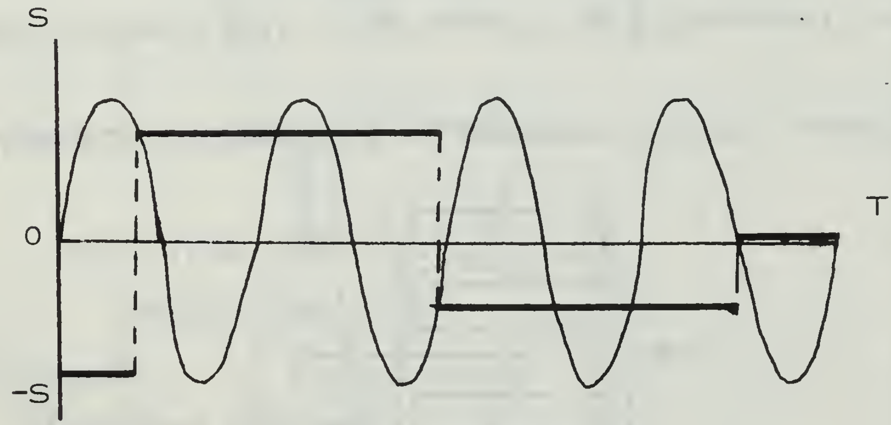
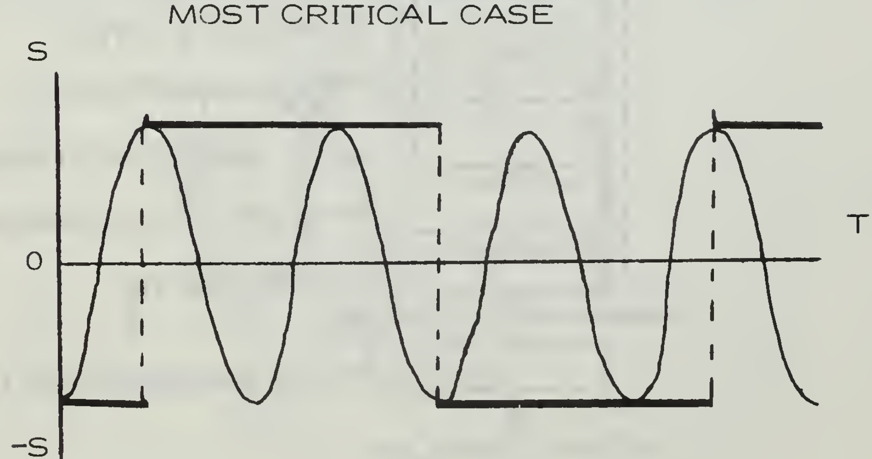


FIGURE 5. SAMPLE-AND-HOLD OUTPUT FOR MOST CRITICAL CASE



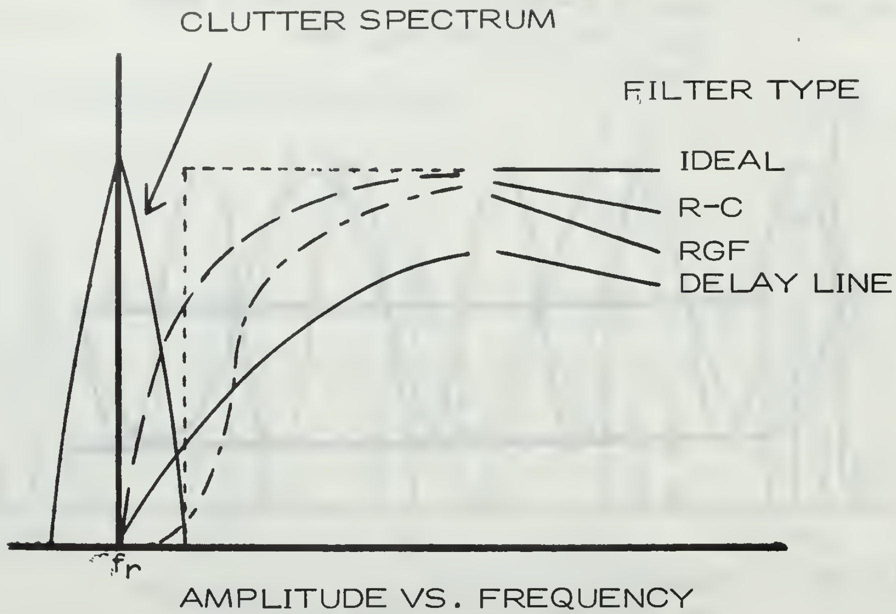
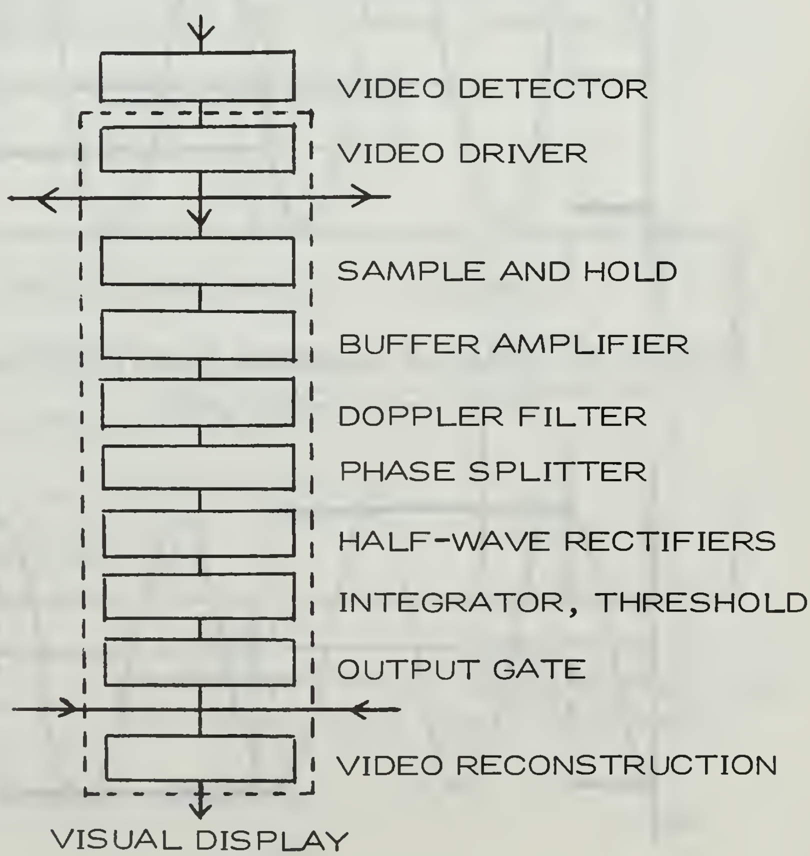


FIGURE 6. LOW-FREQUENCY DOPPLER FILTER CHARACTERISTICS

FIGURE 7. BLOCK DIAGRAM OF SYNTHESIZED CIRCUIT



R in Ohms

C in Microfarads

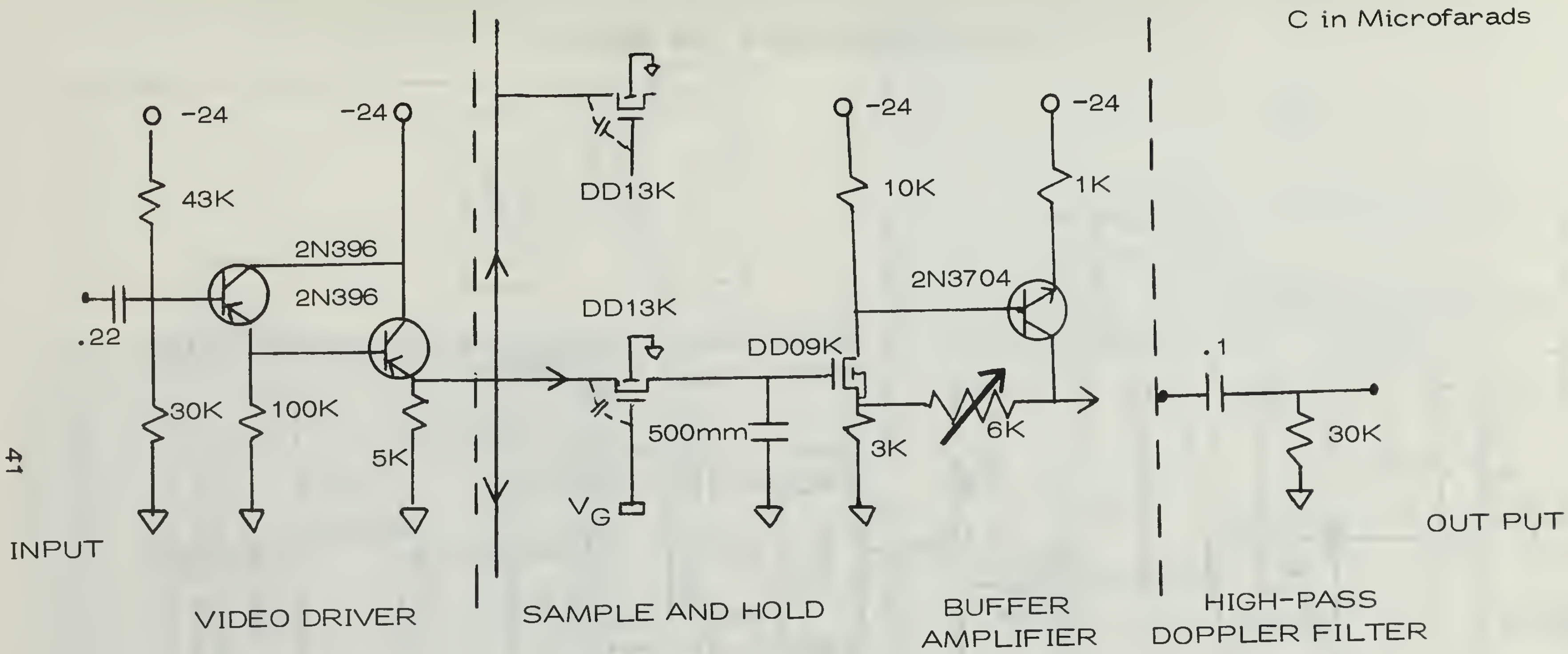


FIGURE 8A: SYNTHESIZED CIRCUIT

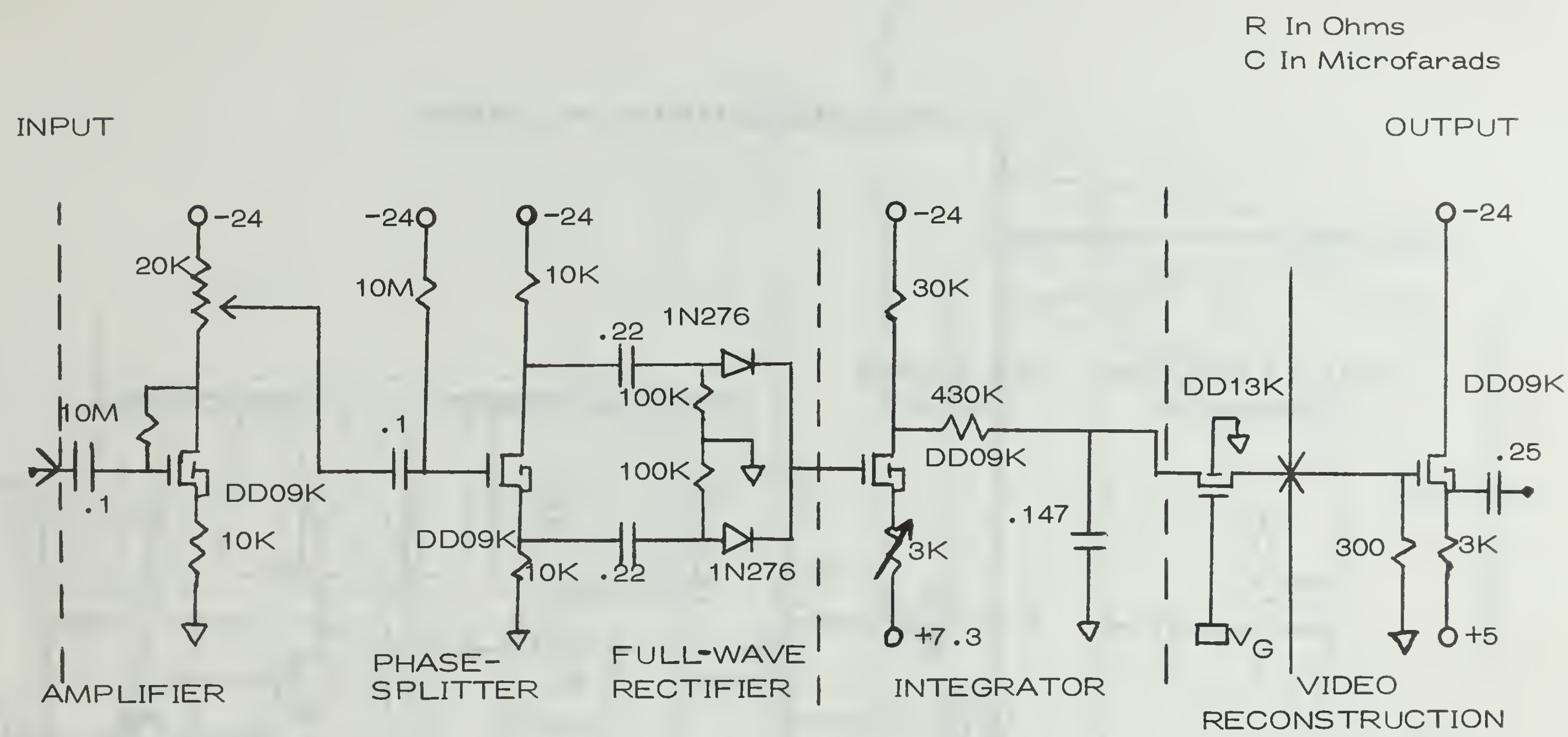


FIGURE 8B: SYNTHESIZED CIRCUIT

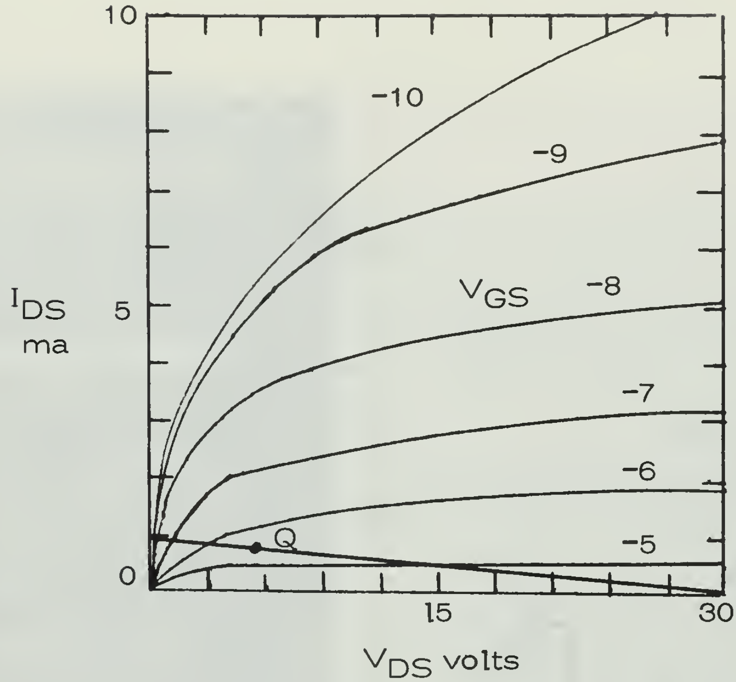
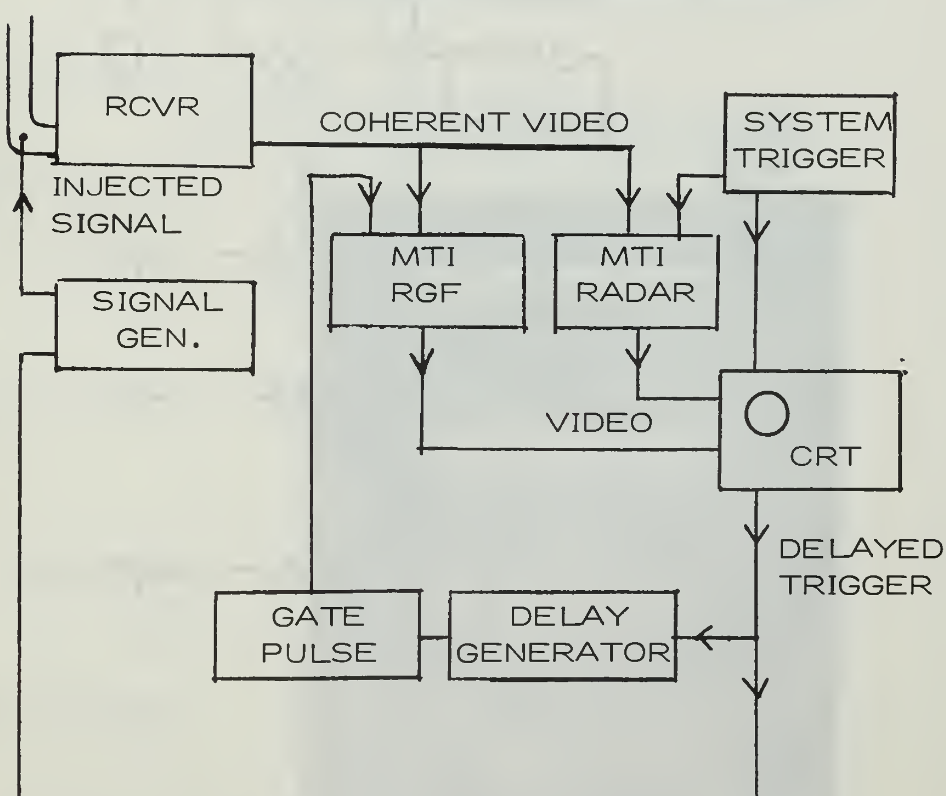


FIGURE 9. CHARACTERISTIC CURVES FOR MOSFET DD09K

FIGURE 10. RADAR-COMPARISON TEST SET-UP



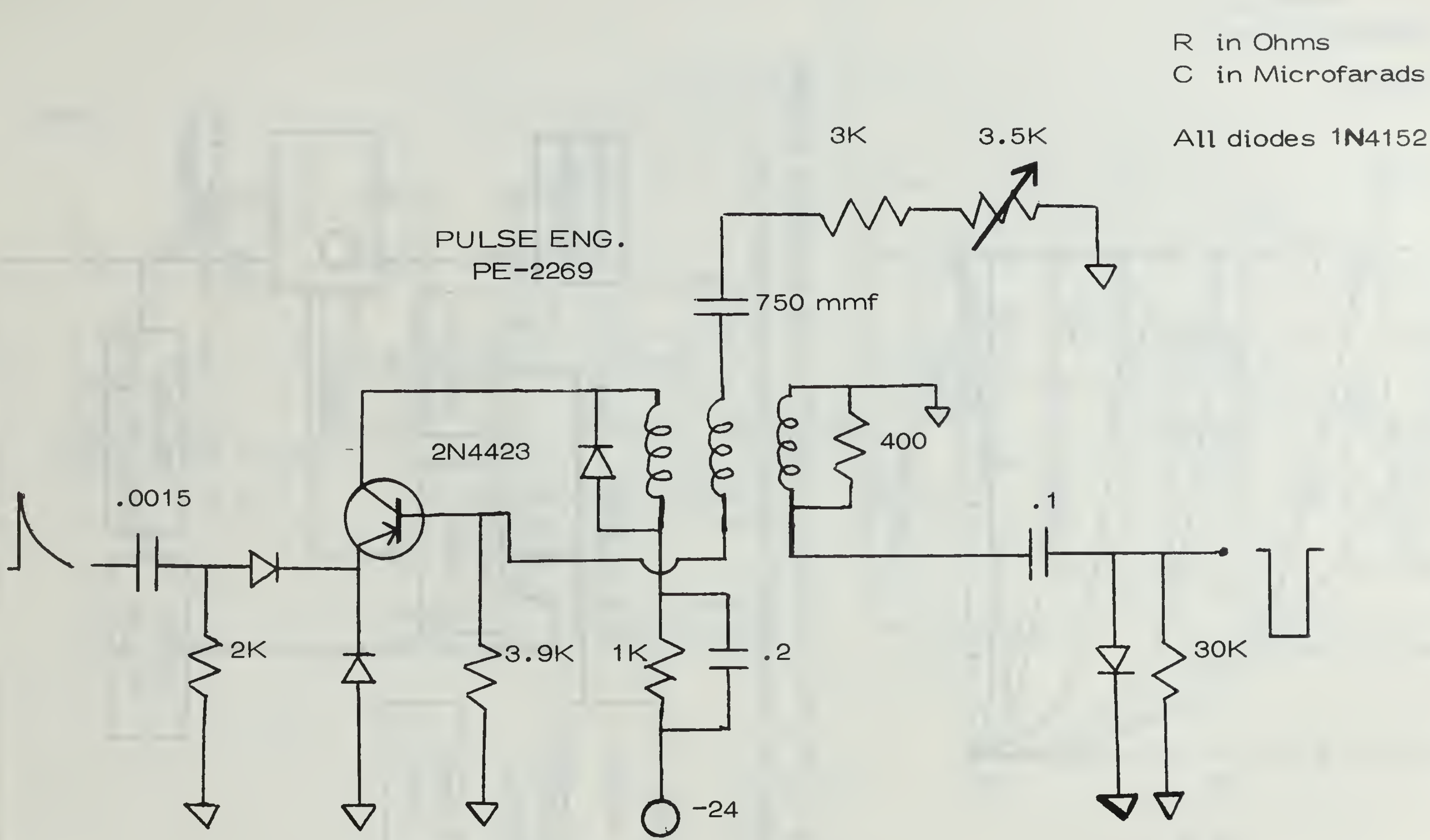


FIGURE 11. RANGE CHANNEL PULSE SOURCE

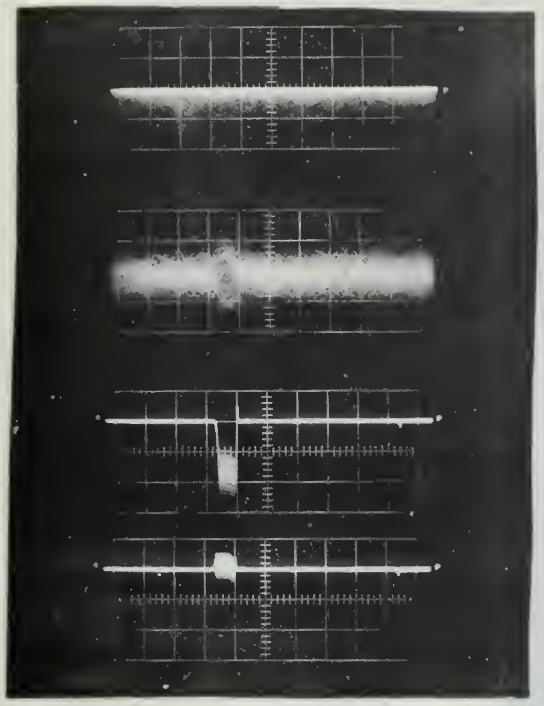
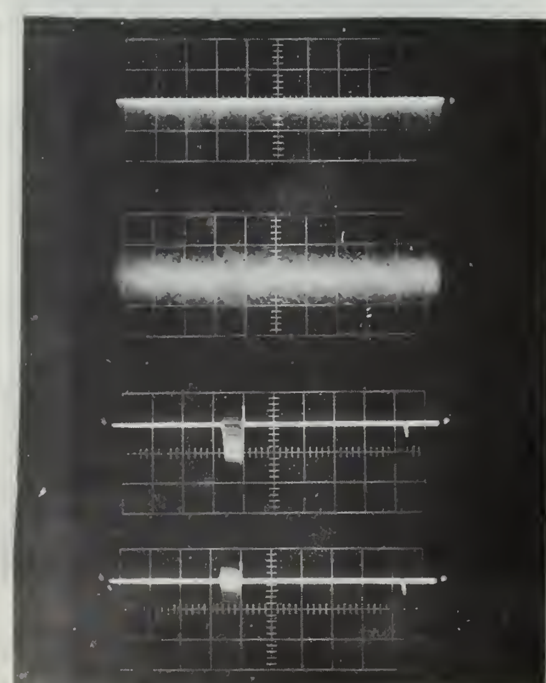
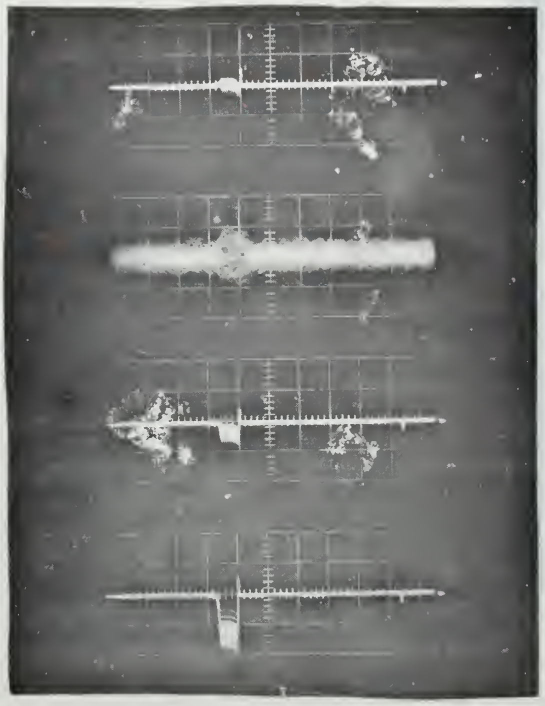


FIGURE 12. COMPARISON OF MTI OUTPUTS FOR A VISIBLE TARGET SIGNAL

FIGURE 13. COMPARISON OF OUTPUTS FOR NON-DISCERNIBLE TARGET SIGNAL





CHANNEL REFERENCE

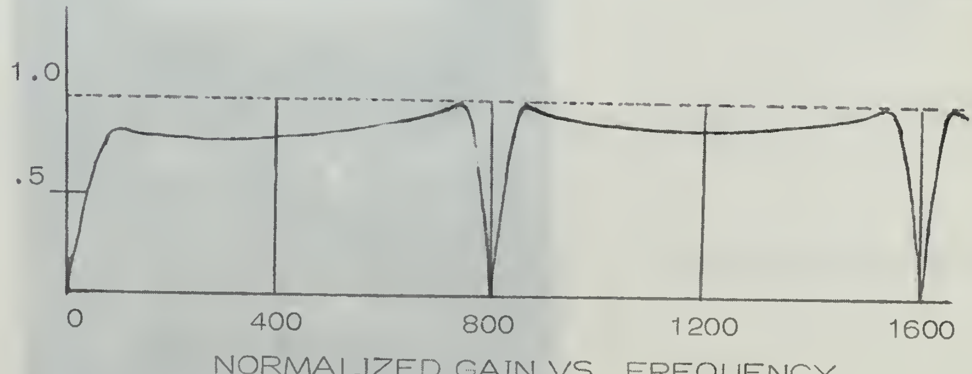
COHERENT VIDEO
(SIGNAL)

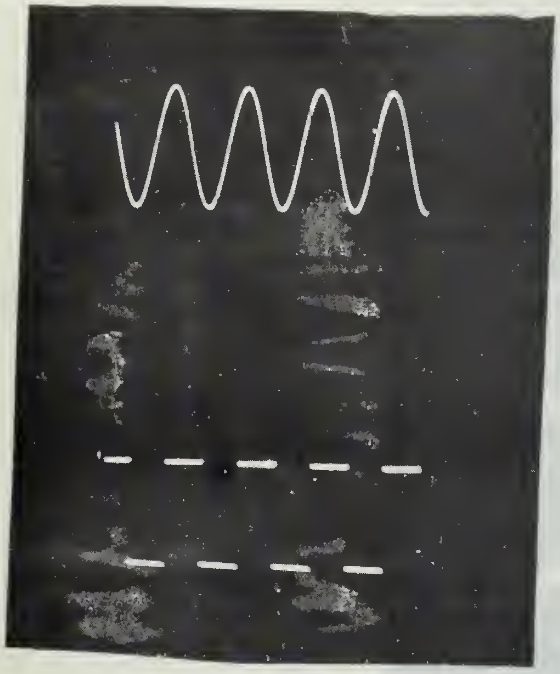
OUTPUT WITH
CAPACITORS

NORMAL CHANNEL
OUTPUT

FIGURE 14. CHANNEL OUTPUT WITH INSERTION
OF LARGE COUPLING CAPACITORS

FIGURE 15. FREQUENCY RESPONSE OF CHANNEL ALONE





SIGNAL

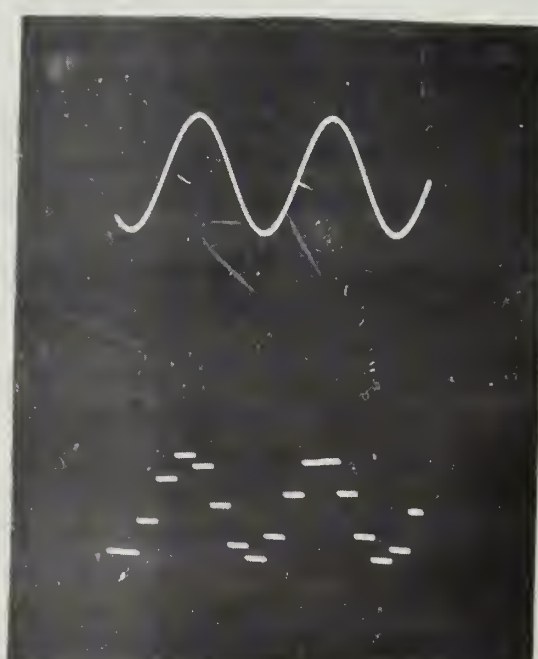
SAMPLES

FIGURE 16. ACTUAL SAMPLE-AND-HOLD OUTPUT FOR 400 CPS SIGNAL

FIGURE 17. ACTUAL SAMPLE-AND-HOLD OUTPUT FOR 110 CPS SIGNAL

SIGNAL

SAMPLES



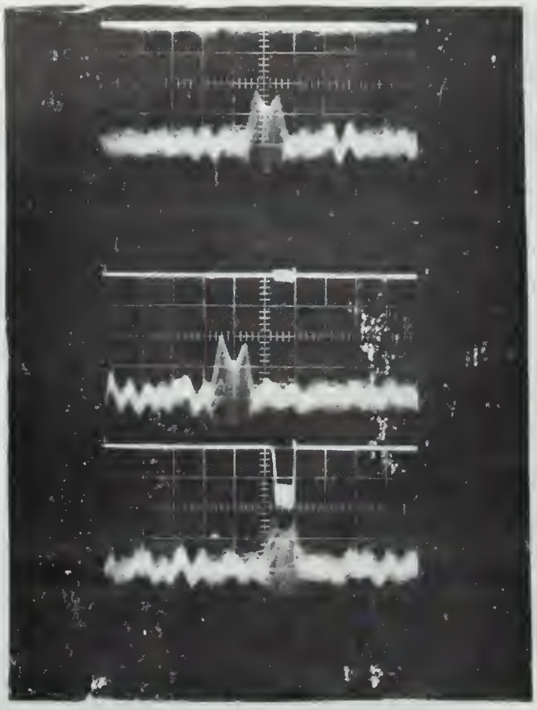


FIGURE 18. COMPARISON OF MTI OUTPUTS FOR AN ACTUAL AIRCRAFT

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Security Classification

DOCUMENT CONTROL DATA - R & D

(Security classification of title, body of abstract and indexing annotation must be entered when the overall report is classified)

1. ORIGINATING ACTIVITY (Corporate author) Naval Postgraduate School Monterey, California 93940	2a. REPORT SECURITY CLASSIFICATION Unclassified 2b. GROUP

3. REPORT TITLE

Synthesis of a Basic Range Channel for Implementation of a Complete MTI by Range-Gated Filtering

4. DESCRIPTIVE NOTES (Type of report and inclusive dates)

Master's Thesis; (June 1969)

5. AUTHOR(S) (First name, middle initial, last name)

Eric L. Washam

6. REPORT DATE	7a. TOTAL NO. OF PAGES	7b. NO. OF REFS
June 1969	50	13

8a. CONTRACT OR GRANT NO. b. PROJECT NO. c. d.	9a. ORIGINATOR'S REPORT NUMBER(S) 9b. OTHER REPORT NO(S) (Any other numbers that may be assigned this report)

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13. ABSTRACT

A range-gated channel was constructed for use with various doppler filters in building the basic element of a radar Moving-Target-Indicator (MTI) by Range-Gated-Filtering (RGF). Improvement over existing systems was accomplished by circuit simplification and solid-state design incorporating MOS devices in sampling circuits and d-c coupled amplifiers. Performance of the channel, using an R-C high-pass filter as the simplest doppler filter, was compared to that of the delay-line canceler MTI of the AN/UPS-1 air-search radar.

DD FORM 1 NOV 65 1473 (PAGE 1)

S/N 0101-807-6811

Security Classification

A-31408

Security Classification

14. KEY WORDS	LINK A		LINK B		LINK C	
	ROLE	WT	ROLE	WT	ROLE	WT
MTI, RANGE-GATED FILTERING, DOPPLER FILTER, SAMPLE-AND-HOLD CIRCUIT, MOSFET, CLUTTER ELIMINATION						

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S/N 0101-807-6821

Security Classification

A-31409